10/578607
IAP20 Rec'd PCT/PTO 08 MAY 2006

METHOD FOR THE HIGHER-ORDER BLIND DEMODULATION OF A LINEAR-WAVEFORM TRANSMITTER

The object of the invention relates to a process for the blind demodulation of signals output by several transmitters and received by an array made up of at least one sensor.

For example, it applies to an array of antennas in an electromagnetic context.

The subject of the invention is in particular the demodulation of signals, that is to say the extraction of the symbols $\{a_k\}$ transmitted by a linearly modulated transmitter.

Figure 1 shows an antenna processing system comprising several transmitters E_i and an antenna processing system T comprising several antennas R_i receiving from radio sources at different angles of incidence. The angles of incidence of the sources or transmitters may be parameterized either in 1D with the azimuth θ_m, or in 2D with the azimuth angle θ_m and the elevation angle Δ_m.

Figure 3 shows schematically a modulation/demodulation principle for the symbols $\{a_k\}$ output by a transmitter. The signal propagates via a multipath channel. The transmitter outputs the symbol a_k at the instant k.T, where T is the symbol period. The demodulation consists in estimating and in detecting the symbols in order to obtain, at the output of the demodulator, the estimated symbols $\hat{a}_k$. In this figure, the train of symbols $\{a_k\}$ is linearly filtered upon transmission by a transmission filter H, also called a wave-shaping filter $h_0(t)$.

In the rest of the description, the expression "blind demodulation" is understood to mean techniques that

basically use no information on the signal output, examples being a wave-shaping filter, a learning sequence, etc.

The last ten years have seen the development of SIMO (single-input, multiple-output) blind demodulation techniques called subspace techniques using 2nd-order statistics, as described in reference [7]. However, these algorithms have the drawback of not being robust to either underestimation or overestimation of the order of the propagation channel, resulting in temporal spreading dependent on the multipaths and on the wave-shaping filter. A linear prediction technique, described in reference [11], has been proposed for overcoming this problem, but this has the drawback of being less effective when the length of the channel is known. To improve the subspace techniques, the method described in [16] proposes a parametric technique, but unfortunately this requires knowledge of the wave-shaping filter.

In reference [13], the authors propose a technique based on covariance matching, but this has in particular the drawback of being very difficult to implement. An easier but suboptimal technique, described in reference [12], was therefore developed by minimizing a likelihood criterion and assuming the symbols to be Gaussian in character. This assumption is not verified for the widely used linear modulations such as PSK (Phase Shift Keying) or QAM (Quadrature Amplitude Modulation).

It is also known, in CMA (Constant Modulus Algorithm) methods, to use a spatio-temporal approach described for example in reference [6]. However, this family of methods has the drawback of being suitable only for one particular class of modulations, such as PSK, which are constant-modulus modulations. This method is iterative and therefore has the drawback of having to be

correctly initialized. Finally, the CMA methods have the disadvantage of converging more slowly than the abovementioned subspace method. Moreover, reference [20] describes a subspace method making use of higher-order statistics for non-minimum-phase FIR (finite impulse response) channel identification.

The subject of the present invention is a process based in particular on blind source separation techniques known to those skilled in the art and described for example in references [4], [5], [15] and [19] assuming that the symbols transmitted are statistically independent. To do this, the process constructs a spatio-temporal observation whose mixed sources are symbol trains from the transmitter. Each symbol train is for example the same symbol train but shifted by an integral number of symbol periods T.

The invention relates to a process for the blind demodulation of a linear-waveform source or transmitter in a system comprising one or more sources and an array of sensors and a propagation channel, said process being characterized in that it comprises at least the following steps:

- the symbol period T is determined and samples are taken at T_e such that $T = IT_e$ (I being an integer);
- a spatio-temporal observation z(t), the mixed sources of which are symbol trains from the transmitter, is constructed from the observations $x(kT_e)$;
- an ICA-type method is applied to the observation vector z(t) in order to estimate the L_c symbol trains $\{a_{m-i}\}$ that are associated with the channel vectors $\hat{h}_{z,j}=\hat{h}_z(k_j)$;
- the L_c outputs $(\hat{a}_{m,j}, \hat{h}_{z,j})$ are arranged in the same order as the inputs $(a_{m-i}, h_z(i))$ so as to obtain the propagation channel vectors $\hat{h}_{z,j}=\hat{h}_z(k_j)$; and
- the phase α_{imax} associated with the outputs is determined.

The process according to the invention offers in particular the following advantages:

- it makes no assumption about the symbol constellations, unlike the methods described in the prior art;
- it requires no knowledge of the wave-shaping filter;
- the modulus of the symbols is not assumed to be constant;
- it is robust to channel length overestimation;
- it can handle propagation channels with correlated paths; and
- it is direct and simple to implement, with no correlated-path crosscheck step.

Other features and advantages of the subject of the present invention will become more clearly apparent on reading the following description, given by way of illustration but implying no limitation, and on examining the appended figures, which show:

- figure 1, an example of an architecture;
- figure 2, the angles of incidence of the sources;
- figure 3, the linear modulation and demodulation process for a symbol train;
- figure 4, the diagram of a linear-modulation transmitter;
- figure 5, a summary of the general principle employed in the invention;
- figure 6, the representation of a constellation;
- figure 7, a first example of the implementation of the method, in which the signal is received baseband;
- figure 8, a second example in which the signal is received in baseband and the multipaths are decorrelated; and
- figure 9, a third example in which the signal is received in baseband and the multipaths are groupwise decorrelated.

To explain the process according to the invention more clearly, the following description relates to a process for the higher-order blind demodulation of a linear-waveform transmitter in an array having a structure as described in figure 1 for example.

Before explaining the steps for implementing the process, the model of the signal used will be described.

Model of the signal output by a source or transmitter
Linear modulation

Figures 3 and 4 show the process for the linear modulation of a symbol train $\{a_k\}$ at the rate T by a wave-shaping filter $h_0(t)$.

The comb of symbols c(t) is firstly filtered by the wave-shaping filter $h_0(t)$ and then transposed to the carrier frequency f_0. The NRZ filter, which is a time window of length T, very often defined by $h_0(t) = \Pi_T(t-T/2)$, is one particular nonlimiting example of a transmission filter. In radio communication, it is also possible to use a Nyquist filter, the Fourier transform of which, $h_0(f) \approx \Pi_B(f-B/2)$, approaches a band window B when the roll-off is zero, therefore $h_0(f) = \Pi_B(f-B/2)$ (the roll-off defines the slope of the filter away from the band B).

The modulates signal $s_0(t)$, output by the transmitter, may be expressed at time $t_k = kT_e$ (T_e being the sampling period) as a function of the comb of symbols c(t):

$$s_0(kT_e) = \sum_i h_0(iT_e)\, c((k-i)T_e) \qquad (1)$$

Let the symbol time T be equal to an integer number of times the sampling period, i.e. $T = IT_e$, and let $k = mI+j$ where $0 \leq j < I$. Since $c(t) = \Sigma_r a_r \delta(t-rIT_e)$, in other words $c(t) = a_u$ for $t = uIT_e$ and $c(t) = 0$ for $t \neq$

uIT_e, the only values for i for which $c((k-i)T_e)$ is non zero satisfy the equation $k-i = uI$, that is to say such that $i = mI+j-uI = nI+j$, where $n = m-u$. Finally, equation (1) becomes:

$$s_0(mIT_e+jT_e)=\sum_{n=-L_0}^{L_0} h_0(nIT_e+jT_e)\,a_{m-n} \quad \text{for} \quad 0 \le j < I \tag{2}$$

The parameter L_0 is the half-length of the transmission filter which is spread over a duration of $(2L_0+1)IT_e$. In the particular case of an NRZ transmission filter, $L_0 = 0$. As regards the transmitted signal $s(t)$, this satisfies the equation $s(t) = s_0(t)\exp(j2\Pi f_0 t)$ as it is equal to the signal $s_0(t)$ transposed to the frequency f_0. Under these conditions, the expression $s(mIT_e+jT_e)$ is, from (2):

$$s(mIT_e+jT_e)=\sum_{n=-L_0}^{L_0} h_0(nIT_e+jT_e)\exp(j2\pi f_0(nI+j)T_e)\,a_{m-n}\exp(j2\pi f_0(m-n)IT_e)$$

$$=\sum_{n=-L_0}^{L_0} h_{F0}(n\,I\,T_e+jT_e)\,b_{m-n} \quad \text{such that} \quad 0 \le j < I \tag{3}$$

where $h_{F0}(iT_e)=h(iT_e)\exp(j2\pi f_0 iT_e)$ and $b_i=a_i\exp(j2\pi f_0 iIT_e)$

Reception of the signals by the sensors

The transmitted signal $s(t)$ (figure 3) passes through a propagation channel before being received on an array made up of N antennas. The propagation channel may be modelled by P multipaths of angle of incidence θ_p, delay τ_p and amplitude ρ_p $(1 \le p \le P)$. At the output of the antennas is the vector $x(t)$, which corresponds to the sum of a linear mixture of P multipaths and noise, assumed to be white and Gaussian. This vector of dimension $N \times 1$ is given by the following expression:

$$x(t)=\sum_{p=1}^{P} \rho_p\,a(\theta_p)\,s(t-\tau_p)+b(t)=A\,s(t)+b(t) \tag{4}$$

where ρ_p is the amplitude of the pth path, $b(t)$ is the noise vector, assumed to be Gaussian, $a(\theta)$ is the response of the array of sensors to a source with angle

of incidence θ, and $A = [a(\theta_1)\ldots a(\theta_p)]$ and $s(t) = [s(t-\tau_1)\ldots s(t-\tau_p)]^T$. Noting that $\tau_p = r_pT+\Delta\tau_p$ (where $0 \leq \Delta\tau_p < T = IT_e$ and r_p is an integer), and inserting equation (3) into equation (4), the vector received by the antennas is given by:

$$x(mIT_e+jT_e) = \sum_{p=1}^{P} \sum_{n=-L_0}^{L_0} \rho_p\, a(\theta_p)\, h_{F0}(n\,IT_e+jT_e-\Delta\tau_p)\, b_{m-n-r_p} + b(mIT_e+jT_e) \qquad (5)$$

By making the change of variable according to $u_p = n+r_p$, the vector received by the antennas is given by:

$$x(mIT_e+jT_e) = \sum_{p=1}^{P} \sum_{u_p=r_p-L_0}^{r_p+L_0} \rho_p\, a(\theta_p)\, h_{F0}((u_p- r_p)IT_e+jT_e-\Delta\tau_p)\, b_{m-u_p} + b(mIT_e+jT_e) \qquad (6)$$

Notating $r_{min} = \min\{r_p\}$ and $r_{max} = \max\{r_p\}$, equation (6) may be rewritten as follows:

$$x(mIT_e+jT_e) = \sum_{p=1}^{P} \sum_{u=r_{min}-L_0}^{r_{max}+L_0} \rho_p\, a(\theta_p)\, h_{F0}((u- r_p)IT_e+jT_e-\Delta\tau_p)\, \mathrm{Ind}_{[rp-L0,\, rp+L0]}(u)\, b_{m-u} + b(mIT_e+jT_e) \qquad (7)$$

where $\mathrm{Ind}_{[r,q]}(u)$ is the usual indicatrix function ($\mathrm{Ind}_{[r,q]}(u) = 1$ for $r \leq u \leq p$ and $\mathrm{Ind}_{[r,q]}(u) = 0$ otherwise) defined over the set of integers relating to the value in the binary set $\{0,1\}$, characterized by $\mathrm{Ind}_{[r,q]}(u) = 1$ if u belongs to the interval $[r,q]$ and $\mathrm{Ind}_{[r,q]}(u) = 0$ otherwise. Thus, denoting the channel vector by $v(t)$:

$$v(uIT_e+jT_e) = \sum_{p=1}^{P} \rho_p\, a(\theta_p)\, h_{F0}((u- r_p)IT_e+jT_e-\Delta\tau_p)\, \mathrm{Ind}_{[rp-L0,\, rp+L0]}(u) \qquad (8)$$

where $t = uIT_e+jT_e$ and equation (5) becomes:

$$x(mIT_e+jT_e) = \sum_{u=r_{min}-L_0}^{r_{max}+L_0} v(uIT_e+jT_e)\, b_{m-u} + b(mIT_e+jT_e) \qquad (9)$$

Inter-symbol interference

The observation vector $x(t)$ coming from the antenna array at $t = mIT_e+jT_e$ involves, from equation (9), the symbol b_m, but also the symbols b_{m-u} where u is a relative integer lying within the $[r_{min}-L_0,\ r_{max}+L_0]$

interval, which phenomenon is more widely known as ISI (inter-symbol interference). Let L_c be the number of symbols participating in the ISI and let the interval of values taken by the latter be limited. From equation (9), if the intersection of the intervals $[r_p-L_0, r_p+L_0]$ is non-empty, then $L_c = |r_{max}-r_{min}| + 2L_0 + 1$. Consequently, when $r_{max} = r_{min}$, that is to say when all the multipaths are correlated, the lower bound of L_c is reached and is given by $L_c = 2L_0 + 1$. This case is also mathematically expressed by $| \max_p\{\tau_p\} - \min_p\{\tau_p\} | < T$. On the other hand, if the intersection of said intervals is empty and if all the intervals $[r_p-L_0, r_p+L_0]$ are disjoint, then $L_c = P\times(2L_0+1)$, which constitutes an upper bound for the set of values that can be taken by L_c. The latter situation corresponds specifically to the case of multipaths that are all pairwise decorrelated. This may also be mathematically expressed as $\forall i \neq j, | r_i - r_j | > 2L_0$, this condition being obtained whenever $| \tau_i - \tau_j | > (2L_0 + 1)T$. To summarize, the quantity L_c is in general bounded as follows:

$$2L_0+1 \leq L_c \leq P\times(2L_0+1) \qquad (10)$$

The equation expressing the vector received by the sensors can then be rewritten in the following manner, but this time only the L_c symbols b_{m-u} of interest appear:

$$X(mlT_e+jT_e) = \sum_{l=1}^{L_c} h(n(l)lT_e+jT_e)\, b_{m-n(l)} + b(mlT_e+jT_e) \qquad (11)$$

where $\forall\ 1 \leq I \leq L_c$ and $r_{min}-L_0 \leq n(I) \leq r_{min}+L_0$ and where:

$$h(t) = \sum_{p=1}^{P} \rho_p\, a(\theta_p)\, h_{F0}(t-\tau_p) \qquad (12)$$

ICA techniques

The process uses ICA techniques based on the following model, given by way of entirely nonlimiting illustration:

$$u_k = \sum_{i=1}^{L} g_i \; s_{ik} + n_k = G\, s_k + n_k \qquad (13)$$

where u_k is a vector of dimension M×1 received at time k, s_{ik} is the ith component of the signal s_k at time k, n is the noise vector and $G = [g_1 \ldots g_L]$. The objective of the ICA methods is to extract the I = L components s_{ik} and to identify their signatures g_i (the vectorial response of source i through the observation u_k) on the basis of the observation u_k. The number I =L of components must not exceed the dimension M of the observation vector. The methods of references [4], [5] and [15] use 2nd- and 4th-order statistics of the observations u_k. The first step uses 2nd-order statistics for the observations u_k (these observations may be functions of the signals received by the sensors) in order to obtain a new observation z_k, such that:

$$z_k = W_1\, u_k = \sum_{i=1}^{L} \breve{g}_i \; s_{ik} + \tilde{n}_k = \breve{G}\, s_k + \tilde{n}_k \qquad (14)$$

where the signatures $\breve{g}_i$ ($1 \le i \le L$) are orthogonal, $\breve{G} = [\breve{g}_1 \ldots \breve{g}_L]$ and $s_k = [s_{1k} \ldots s_{Lk}]^T$. The second step consists in identifying the orthogonal base of the $\breve{G}$ values using 4th order statistics of the whitened observations z_k. In this way, the signals s_k may be extracted by effecting:

$$\hat{s}_k = \breve{G}^{\#} z_k = \breve{G}^{\#} W_1\, u_k \qquad (15)$$

where $\hat{s}_k$ is the estimate of the signals s_k and where $^{\#}$ is the pseudo-inversion operator defined by $\breve{G}^{\#} = (\breve{G}^H \breve{G})^{-1} \breve{G}^H$.

The ICAR method [19] uses only 4th-order statistics to identify the matrix $G = [g_1 \ldots g_k]$ of signatures.

To summarize, the idea employed in the process according to the invention is to construct a spatio-temporal observation, the mixed sources of which are symbol trains from the transmitter. Each symbol train

is for example the same symbol train but shifted by an integral number of symbol periods T.

Several ways of implementing the method will be described below, some of which are explained by way of non-limiting illustration.

First way of implementing the process.

Figure 7 shows a first illustrative way of implementing the process in which the signal is received in baseband.

The method comprises a step I.1 of determining the symbol time T_e, for example by applying a cyclic detection algorithm, such as that described for example in [1] [10].

The next step I.2 consists in interpolating the observations x(t) with I samples per symbol, such that $T = IT_e$.
Under these conditions where $f_0 = 0$ and $b_k = a_k$, equation (11) for the vector becomes:

$$x(ml\,T_e + jT_e) = \sum_{l=1}^{L_c} h(n(l)\,I\,T_e + jT_e)\, a_{m-n(l)} + b(ml\,T_e + jT_e) \quad \text{for} \quad 0 \leq j < I \qquad (16)$$

Since equation (16) is valid for $0 \leq j < I$, the method constructs the next spatio-temporal observation (step I3) from the observations $x(kT_e)$:

$$z(ml\,T_e) = \begin{bmatrix} x(mIT_e) \\ x(mIT_e + T_e) \\ \vdots \\ x(mIT_e + (I-1)T_e) \end{bmatrix} = \sum_{l=1}^{L_c} h_z(n(l))\, a_{m-n(l)} + b_z(ml\,T_e) \text{ where } h_z(n) = \begin{bmatrix} \mathbf{h}_{n,0} \\ \mathbf{h}_{n,1} \\ \vdots \\ \mathbf{h}_{n,I-1} \end{bmatrix} \qquad (17)$$

with $h_{n,j} = h(nI\,T_e + jT_e)$ and $b_z(ml\,T_e) = [\,b(ml\,T_e)^T \ldots b(ml\,T_e + (I-1)T_e)^T]^T$.

Since it is known that x(t) has the dimensions N×1, the vector z(t) has the dimensions NI×1.

$h(k)$ is a vector whose nth component is the kth component of the filter that linearly filters the symbol train $\{a_m\}$ on the nth sensor. The filter for the vector coefficient $h(k)$ depends both on the wave-shaping filter and on the propagation channel.

To extract the L_c symbol trains $\{a_{m-i}\}$ of interest (the number of symbols participating in the ISI), the method samples the received signal with $I = (2L_0+1)$, assuming that $P \leq N$.

Since it is known that the NRZ filter satisfies $2L_0+1 = 1$ and the Nyquist filter $2L_0+1 = 3$ for a roll-off of 0.25, the symbol trains may be extracted for these two wave-shaping filters when $P \leq NI$ and $3P \leq NI$, respectively.

Having determined the observation vector $z(t)$, the process applies an ICA-type method to estimate the L_c symbol trains $\{a_{m-i}\}$ associated with the channel vectors $\hat{h}_{z,j} = \hat{h}_z(k_j)$.

The jth output of the ICA method gives the symbol train $\{\hat{a}_{m,j}\}$ associated with the channel vector $\hat{h}_{z,j}$. The estimated symbol trains $\{\hat{a}_{m,j}\}$ arrive in a different order from that of the $\{a_{m-i}\}$ trains satisfying:

$$\hat{a}_{m,j} = \rho \exp(j\alpha_i)\, a_{m-i} \text{ and } \hat{h}_{z,j} = h_z(i) \tag{18}$$

The symbol trains $\{\hat{a}_{m,j}\}$ are estimated with the same amplitude because the symbol trains $\{a_{m-i}\}$ all have the same power, satisfying the equation:

$$E[|a_{m-n(1)}|^2] = \ldots = E[|a_{m-n(Lc)}|^2].$$

The next step I.4 of the process has the objective of ordering the L_c outputs $(\hat{a}_{m,j}, \hat{h}_{z,j})$ in the same order as the inputs $(a_{m-i}, h_z(i))$ so as to obtain the channel vectors $\hat{h}_{z,j} = \hat{h}_z(k_j)$. To do this, the process

intercorrelates pairwise the outputs $\hat{a}_{m,i}$ and $\hat{a}_{m,j}$, calculating the following criterion $c_{i,j}(k)$:

$$c_{i,j}(k) = \frac{E[\hat{a}_{m,i}\,\hat{a}_{m-k,j}{}^*]}{\sqrt{E[\hat{a}_{m,i}\hat{a}_{m,i}{}^*]E[\hat{a}_{m-k,j}\hat{a}_{m-k,j}{}^*]}} \qquad (19)$$

When the function $|c_{i,j}(k)|$ is a maximum in $k = k_{max}$, the ith and jth outputs satisfy the equation: $\hat{a}_{m,i} = \hat{a}_{m-k\,max,j}$. The algorithm for classifying the outputs $\hat{a}_{m,n(1)}\ldots\hat{a}_{m,n(Lc)}$ is for example composed of the following steps:

Step A.1: Determination of the output $\hat{a}_{m,imax}$ associated with the channel vector of higher-modulus $\hat{\mathbf{h}}_{z,jmax}$.

Step A.2: For all the outputs $\hat{a}_{m-k,j}$ where $j \neq i_{max}$, determination of the indices $k = k_j$ maximizing the $|c_{imax,j(k)}|$ criterion. From this is deduced, for each j, that $\hat{a}_{m,imax} = \hat{a}_{m-kj,j}$. Since it is known that $c_{imax,j}(k_j) = \exp(j\alpha_{imax}-j\alpha_j)$ the jth output is reset to the same phase as the iith i_{max} output by taking $\hat{a}_{m-kj} = c_{imax,j}(k_j)\,\hat{a}_{m,j}$. The channel vectors are also reset in terms of phase by taking: $\hat{\hat{\mathbf{h}}}_z(k_j) = \hat{\mathbf{h}}_{z,j}\,c_{imax,j}(k_j)^*$.

Step A.3: This step reorders the outputs $\hat{a}_{m-kj}$ and the channel vectors $\hat{\hat{\mathbf{h}}}_{z,j} = \hat{\mathbf{h}}_z(k_j)$ in the increasing order of the K_j, since it is known that $\hat{a}_m = \hat{a}_{m,imax}$ and that $\hat{\hat{\mathbf{h}}}_z(0) = \hat{\mathbf{h}}_{z,imax}$.

After these three steps, the symbol trains $\{\hat{a}_{m-k}\}$ associated with the channel vectors $\hat{\mathbf{h}}_z(k_j)$ are obtained. Since it is known that the estimated symbols satisfy the equation $\hat{a}_{m-k} = \exp(j\alpha_{imax})a_{m-k}$, the last step of the process consists in estimating this phase α_{imax}. To do this, the constellation of symbols a_k is firstly identified among a database consisting of the set of possible constellations. This database consists of known constellations such as nPSK, n-QAM. Each time that a new constellation is detected or becomes known, this is added to the database.

Figure 6 shows an example of an 8-QAM constellation when $\alpha_{imax}=0$ and $\alpha_{imax}\neq 0$. In this

implementation example, the process then includes the following steps:

The next step I.5 consists in determining the output phase associated with the channel vector of higher modulus. To identify the constellation and determine the phase, the process performs, for example, the following steps:

Step I.5 = Steps B.1, B.2 and B.3

Step°B.1: Estimation of the positions of the states of the constellation (red points in the figure) by seeking the maximum of the 2D histogramme of the points $M_k=(\mathrm{real}(\hat{a}_k), \mathrm{imag}(\hat{a}_k))$. For a constellation consisting of M states, M pairs $(\hat{u}_m, \hat{v}_m)$ for $1 \leq m \leq M$ are obtained.

Step°B.2 : Determination of the type of constellation by comparing the position of the states $(\hat{u}_m, \hat{v}_m)$ of the constellation of $\{\hat{a}_k\}$ symbols with a database comprising the set of possible constellations. The closest constellation is made up of the states (u_m, v_m) for $1 \leq m \leq M$.

Step B.3: Determination of the phase α_{imax} by minimizing in the sense of the least squares the following system of equations :

$$\hat{u}_m=\cos(\alpha_{imax})u_m - \sin(\alpha_{imax})v_m$$

and $\hat{v}_m = \sin(\alpha_{imax})u_m + \cos(\alpha_{imax})v_m$ for $1 \leq m \leq M$.

The process may include a step of estimating the propagation channel parameters of angle θ_p and delay τ_p, of equation (8) by the algorithm proposed in [8]. The step consists in extracting firstly the vectors $h(nIT_e+ jT_e)$ for $0 \leq j < I$ from the channel vectors $\hat{\mathbf{h}}_z(n_j)$ defined in equation (17). Followed by construction of the matrix $H=[h(n(1)IT_e)\ldots h(n(L_c)IT_e)]$ from equation (11) with the $h(nIT_e+ jT_e)$ values in order to apply the parametric estimation method [8] for the multipaths: (θ_p, τ_p) $1 \leq p \leq P$.

Second way of implementing the process

Figures 8 and 9 show schematically another way of implementing the process, which may include two variants corresponding to the decorrelated multipath case and to the groupwise correlated multipath case, respectively.

Decorrelated multipath case.

The signal is received in baseband with $\{b_k\}=\{a_k\}$.

The multipaths, the delays of which satisfy the relationship $|\tau_j-\tau_i| > (2L_0+1)T$, have the advantage of being decorrelated with one another, satisfying the equation : $E[s(t-\tau_i)s(t-\tau_j)^*] = 0$. By examining equation (4), it may therefore be seen that it is sufficient to apply an ICA type method when $P \leq N$ to the observation $x(t)$ in order to obtain the signals $s(t-\tau_p)$ for each of the multibars. After estimating the signals for the various multipaths, the process determines their powers in order to keep the signal $s(t-\tau_{pmax})$ of the multipath of higher amplitude ρ_{pmax}. This main path is determined using the fact that the outputs of the ICA methods asymptotically satisfy:

$$x(t)=\sum_{p=1}^{P} \rho_p a(\theta_p) s(t-\tau_p) = \sum_{i=1}^{P} \hat{a}_i \hat{s}_i(t) \quad \text{with} \qquad (20)$$

$$\hat{s}_i(t) = \frac{s(t-\tau_p)}{\sqrt{\gamma_p}} \quad \text{and} \quad \hat{a}_i = \sqrt{\gamma_p}\, \rho_p\, a(\theta_p)$$

where $\gamma_p=\rho_p^2 E[|s(t-\tau_p)|^2]$. Since the vectors $a(\theta_p)$ are normed, satisfying the equation $a(\theta_p)^H a(\theta_p)=N$, the path of maximum amplitude will be associated with the i_{max}^th output where $\alpha_{imax}=\hat{a}_{imax}^H \hat{a}_{imax}$ is a maximum. From equation (3), the output $\hat{s}_{imax}(t)=s(t-\tau_{pmax})$ satisfies the equation :

$$\hat{s}_{imax}(mIT_e+jT_e)=\sum_{n=-L_0}^{L_0} h_{F0}(nIT_e+jT_e-\tau_{pmax})\, a_{m-n} \qquad (21)$$

such that $0 \leq j < I$

and it is possible to constitute the following observation vector:

$$z(mIT_e)=\begin{bmatrix} \hat{s}_{i\max}(mIT_e) \\ \hat{s}_{i\max}(mIT_e+T_e) \\ \vdots \\ \hat{s}_{i\max}(mIT_e+(I-1)T_e) \end{bmatrix} = \sum_{n=-L_0}^{L_0} h_z(n)\, a_{m-n} \tag{22}$$

where $h_z(n)=\begin{bmatrix} h_{n,0} \\ h_{n,1} \\ \vdots \\ h_{n,I-1} \end{bmatrix}$ and

where $h_{n,j}=h_{F0}(nIT_e+jT_e-\tau_{pmax})$. According to the model of equation (22), it is sufficient to apply an ICA method to the observation $z(mIT_e)$ in order to estimate the $2L_0+1$ symbol trains $\{a_{m-n}\}$ with $-L_0 \leq n \leq L_0$. To extract the angles of incidence θ_p of the propagation channel, it is sufficient from equation (20) to find, for each signature $\hat{a}_i$ $(1 \leq i \leq P)$, the maximum of criterion $c(\theta)=|a(\theta)^H\hat{a}_i|^2$. To extract the delays $\tau_i - \tau_1$ of the propagation channel, it is sufficient from equation (20) to find, for each signal $\hat{s}_i(t)$ $(1 \leq i \leq P)$, the maximum of the $c(\tau)=|\hat{s}_i(t-\tau)\,\hat{s}_1(t)^*|^2$ criterion.

To summarize, this variant comprises, for example, the following steps:

Step II.a.1: Determination of the symbol period T, applying a cyclic detection algorithm as in [1][10].

Step II.a.2: Sampling of the observations x(t) with I samples per symbol such that $T = I\,T_e$.

Step II.a.3: Application of an ICA method to the observations x(t) in order to obtain $\hat{s}_i(t)$ and $\hat{a}_i$ for $1 \leq i \leq P$.

Step II.a.4: Determination of the output i = imax where $\alpha_i=\hat{a}_i^H\hat{a}_i$ is its maximum.

Step II.a.5: Formation of the observation vector z(t) of equation (22) from the signal $\hat{s}_{i\max}(t)$.

Step II.a.6: Application of an ICA method for

estimating the symbol trains $\{a_{m-n}\}$ where $-L_0 \leq n \leq L_0$. From the symbol trains is chosen that one which is associated with the higher-modulus vector $h_z(n)$, namely $\{\hat{a}_m\}$.

Step II.a.7: Determination of the phase α_{imax} of the output associated with the higher-modulus vector $h_z(n)$ applying steps B.1, B.2 et B.3.

Step II.a.8: Phase-resetting of the symbol train $\{\hat{a}_m\}$ by taking $\hat{\hat{a}}_m = \hat{a}_m \exp(-j\alpha_{imax})$. The symbol train $\{\hat{\hat{a}}_m\}$ constitutes the output of the demodulator of this subprocess.

Step II.a.9: Estimation of the propagation channel parameters, namely angle θ_p and delay τ_p, by maximizing, for $1 \leq i \leq P$ the $|a(\theta)^H \hat{a}_i|^2$ and $|\hat{s}_i(t-\tau)\, \hat{s}_1(t)^*|^2$ criteria for the angles and delays respectively.

General case for any or groupwise-correlated multipaths

In this variant, the diagram for which is given in figure 9, the process considers that some of the multipaths are correlated. Considering that the transmitter is received according to Q groups of correlated multipaths, the signal vector received by the sensors becomes, from equation (4):

$$x(t) = \sum_{q=1}^{Q} \sum_{p=1}^{P_q} \rho_{p,q}\, a(\theta_{p,q})\, s(t-\tau_{p,q}) + b(t) \tag{23}$$

$$= \sum_{q=1}^{Q} A_q\, \Omega_q\, s(t, \underline{\tau}_q) + b(t)$$

Where $A_q = [a(\theta_{1,q}) \ldots a(\theta_{Pq,q})]$, $\Omega_q = \mathrm{diag}([\rho_{1,q} \ldots \rho_{Pq,q}])$ and $s(t, \underline{\tau}_q) = [s(t-\tau_{1,q}) \ldots s(t-\tau_{Pq,q})]^T$ with $\underline{\tau}_q = [\tau_{1,q} \ldots \tau_{Pq,q}]^T$. The following signals and signatures are estimated as output of the separator by applying an ICA method:

$$\hat{A}=[\hat{a}_1\ldots\hat{a}_{PQ,Q}]=[A_1U_1\ldots A_QU_Q]\,\Pi \text{ and } \hat{s}(t)=\Pi\begin{bmatrix} \mathbf{V}_1\, s(t,\underline{\tau}_1) \\ \vdots \\ \mathbf{V}_Q\, s(t,\underline{\tau}_Q) \end{bmatrix}=\begin{bmatrix} \hat{s}_1(t) \\ \vdots \\ \hat{s}_{P_Q x Q}(t) \end{bmatrix} \quad (24)$$

where Π is a permutation matrix, $U_qV_q = \Omega_q$ and $V_qE[s(t,\underline{\tau}_q)s(t,\underline{\tau}_q)^H]V_q^H=I_{Pq}$. Thus, the paths decorrelated such that $E[s(t-\tau_{p,q})s(t-\tau_{p',q'})^*]=0$ are received on different channels $\hat{s}_i(t)$ and $\hat{s}_j(t)$. The correlated paths where $E[s(t-\tau_{p,q})s(t-\tau_{p',q'})^*]\neq 0$ are mixed in the same channel $\hat{s}_i(t)$ and are present on P_Q at the same time. In the 1st step of this subprocess, we use this result to identify the Q group of correlated multipaths. Taking the outputs i and j of the separator, the two following hypotheses may be tested:

$$H_0: \begin{cases} \hat{s}_i(t)=b_i(t) \\ \hat{s}_j(t)=b_j(t) \end{cases} \text{ and } H_1: \begin{cases} \hat{s}_i(t)=\alpha_i\, s(t-\tau_p)+b_i(t) \\ \hat{s}_j(t)=\alpha_j\, s(t-\tau_p)+b_j(t) \end{cases} \quad (25)$$

where $E[b_i(t)\ b_j(t-\tau)^*] = 0$ whatever the value of τ. Thus for the H_0 hypothesis, no multipaths exist common to the two output i and j, and for the H_1 hypothesis there is at least one of them. The test consists in determining whether the outputs $\hat{s}_i(t)$ and $\hat{s}_j(t-\tau)$ are correlated for at least one of the τ values satisfying $|\tau| < \tau_{max}$. To do this, the Gardner test [3] is applied, which compares the following likelihood ratio with a threshold:

$$V_{ij}(\tau)=-2K\ln\left(1-\frac{|\hat{r}_{ij}(\tau)|^2}{\hat{r}_{ii}(0)\hat{r}_{jj}(0)}\right) \text{ with } \hat{r}_{ij}(\tau)=\frac{1}{K}\sum_{k=1}^{K}\hat{s}_i(t)\hat{s}_j(t-\tau)^* \quad (26)$$

or $V_{ij}(\tau)<\eta \Rightarrow$ hypothesis H_0

And $V_{ij}(\tau)\geq\eta \Rightarrow$ hypothesis H_1

The threshold η is determined in [3] in relation to a chi square law with 2 degrees of freedom. The output

associated with the 1st output are firstly sought by starting the test by $2 < j \leq P_Q \times Q$ and $i = 1$. Next, removed from the list of outputs are all those associated with the 1st which will constitute the 1st group with $q = 1$. The same series of tests is then restarted with the other outputs not correlated with the 1st output in order to constitute the 2nd group. This operation will be carried as far as the last group where, in the end, no output channel will remain. After the sorting, what will finally be obtained are:

$$\hat{A}_q = A_q U_q \text{ and } \hat{s}_q(t) = V_q s(t, \underline{\tau}_q) \text{ for } (1 \leq q \leq Q) \qquad (27)$$

The angles of incidence $\theta_{p,q}$ are determined from the $\hat{A}_q$ values for $(1 \leq q \leq Q)$ applying the MUSIC [1] algorithm to the $\hat{A}_q\hat{A}_q^H$ matrix. The matrices A_q are deduced from these goniometry values. Since is it known that $x_q(t) = \hat{A}_q\ \hat{s}_q(t) = A_q\ \Omega_q\ s(t, \underline{\tau}_q)$, $s(t, \underline{\tau}_q)$ is deduced therefrom to within a diagonal matrix by taking $\hat{s}(t, \underline{\tau}_q) = A_q^{\#} x_q(t)$. Since the elements of $\hat{s}(t, \underline{\tau}_q)$ are composed of the signals $\hat{s}(t-\tau_{p,q})$, the delays $\tau_{p,q}-\tau_{1,1}$ are determined by maximizing the $c(\tau) = |\hat{s}_{q,p}(t-\tau)\hat{s}_{1,1}(t)^*|^2$ criteria where $\hat{s}_{q,p}(t)$ is the p^{th} component of $\hat{s}(t, \underline{\tau}_q)$.

Since it is known that $E[\hat{s}_q(t)\hat{s}_q(t)^H] = I_{Pq}$, that $A_q^H A_q = N\ I_{Pq}$ and that $\hat{A}_q \hat{s}_q(t) = A_q \Omega_q s(t, \underline{\tau}_q)$, it is deduced therefrom that the group of multipaths associated with the largest amplitudes Ω_q maximizes the following criterion: $cri(q) = \text{trace}(\hat{A}_q^H\ \hat{A}_q)$. From this is deduced the best output associated with $\hat{A}_{qmax}$ and $\hat{s}_{qmax}(t)$. Since from equation (3) the vector $s(t, \underline{\tau}_{qmax})$ satisfies equation:

$$s(mIT_e+jT_e, \underline{\tau}_{qmax}) = \begin{bmatrix} s(mIT_e + jT_e - \tau_{q\max 1}) \\ \vdots \\ s(mIT_e + jT_e - \tau_{q\max P_{q\max}}) \end{bmatrix} = \sum_{n=-L_0}^{L_0} h_{F0}(nIT_e+jT_e, \underline{\tau}_{qmax})\, a_{m-n} \qquad (28)$$

for

$0 \le j < I$ and with $h_{F0}(nIT_e+jT_e,\underline{\tau}_{qmax})=\begin{bmatrix} h_{F0}(nIT_e+jT_e-\tau_{q\max,1}) \\ \vdots \\ h_{F0}(nIT_e+jT_e-\tau_{q\max,P_{q\max}}) \end{bmatrix}$

it is possible to constitute the following observation vector from equation (27):

$$z(mIT_e)=\begin{bmatrix} \hat{\mathbf{s}}_{q\max}(mIT_e) \\ \hat{\mathbf{s}}_{q\max}(mIT_e+T_e) \\ \vdots \\ \hat{\mathbf{s}}_{q\max}(mIT_e+(I-1)T_e) \end{bmatrix} = \sum_{n=-L_0}^{L_0} h_z(n)\, a_{m-n} \text{ where } h_z(n)=\begin{bmatrix} \mathbf{h}_{n,0} \\ \mathbf{h}_{n,1} \\ \vdots \\ \mathbf{h}_{n,I-1} \end{bmatrix} \quad (29)$$

where $h_{n,j} = V_{qmax}\, h_{F0}(nIT_e + jT_e, \underline{\tau}_{qmax})$. From the model of equation (29), it is sufficient to apply an ICA method to the observation $z(mIT_e)$ in order to estimate the $2L_0 + 1$ symbol trains $\{a_{m-n}\}$ such that $-L_0 \le n \le L_0$.

To summarize, this variant comprises the following steps:

Step II.b.1: Determination of the symbol period T by applying a cyclic detection algorithm as in [1] and [10].

Step II.b.2: Sampling of the observations x(t) with I samples per symbol such that $T=IT_e$.

Step II.b.3: Application of an ICA method to the observations x(t) in order to obtain $\hat{\mathbf{s}}(t)$ and $\hat{A}$ from equation (24).

Step II.b.4: Sorting of the outputs according to Q groups of correlated multipaths in order to obtain $\hat{A}_q$ and $\hat{s}_q(t)$ for $(1 \le q \le Q)$: to do this, a correlation test for all the output pairs i and j with the two-hypothesis test of equation (26). Firstly the outputs associated with the 1st output will be sought by starting the test for $2 < j \le P_Q \times Q$ and $i = 1$. Next, removed from the list of outputs are all those associated with the 1st that will constitute the 1st group with $q = 1$. The same series of tests is repeated

with the other outputs that are not correlated with the 1st output in order to constitute the 2nd group. This operation is continued to the last group where in the end no output channel will remain.

Step II.b.5: Determination of the better group of multipaths where $(\hat{A}_q^H \hat{A}q)$ is a maximum in $q = qmax$.

Step II.b.6: Constitution of the observation vector z(t) of equation (29) from the signal $\hat{s}_{q\,max}(t)$.

Step II.b.7: Application of an ICA method for estimating the symbol trains $\{a_{m-n}\}$ where $-L_0 \leq n \leq L_0$. From the symbol trains is chosen that one which is associated with the higher-modulus vector $h_z(i)$, namely $\{\hat{a}_{m-i}\}$.

Step II.b.8: Determination of the phase α_{imax} of the output associated with the higher-modulus vector $h_z(i)$ applying steps B.1, B.2 and B.3.

Step II.b.9: Phase-resetting of the symbol trains $\{\hat{a}_m\}$ by taking $\hat{\hat{a}}_m = \hat{a}_m \exp(-j\alpha_{imax})$. The symbol train $\{\hat{\hat{a}}_m\}$ constitutes the output of the demodulator of this subprocess.

Step II.b.10: Estimation of the propagation channel parameters, namely the angle $\theta_{q,p}$ and the delay $\tau_{q,p}$. The angles of incidence $\theta_{q,p}$ are determined from the $\hat{A}_q$ values for $(1 \leq q \leq Q)$ applying the MUSIC [1] algorithm to the matrix $\hat{A}_q \hat{A}_q^H$. The matrices A_q are deduced from these goniometry values in order to deduce therefrom an estimate of $s(t, \underline{\tau}_q)$ taking $\hat{s}(t, \underline{\tau}_q) = A_q^{\#} x_q(t)$. Since the elements of the $\hat{s}(t, \underline{\tau}_q)$ are composed of the signals $s(t-\tau_{q,p})$, the delays $\tau_{p,q}-\tau_{1,1}$ are determined by maximizing the $c(\tau)=|\hat{s}_{q,p}(t-\tau)\hat{s}_{1,1}(t)^*|^2$ critera where $\hat{s}_{q,p}(t)$ is the p^{th} component of $\hat{s}(t, \underline{\tau}_q)$.

Another way of implementing the process

Estimation of the carrier frequency and deduction of the $\{a_m\}$ symbols.

This technique consists in estimating the carrier frequency f_0 of the transmitter or the complex $z_0=\exp(j2\pi f_0 T_e)$ in order thereafter to deduce the

symbols $\{a_m\}$ from the symbols $\{b_m\}$, taking, from equation (3):

$$a_m=b_m \exp(-j2\pi f_0 mIT_e)=b_m z_0^{-mI} \qquad (30)$$

This step is applied after step I.4 of reordering the symbols and the channel vectors. From equations (3), (17), (7) and (8), the following channel vectors are used:

$$\hat{\mathbf{h}}_z(n) = \begin{bmatrix} z_0^{nI}\,\mathbf{h}(nIT_e) \\ z_0^{nI+1}\,\mathbf{h}(nIT_e+T_e) \\ \vdots \\ z_0^{nI+(I-1)}\mathbf{h}(nIT_e+(I-1)T_e) \end{bmatrix} \text{ for } n\in\Omega \qquad (31)$$

where $\Omega=\{\mathrm{Ind}_{[rp-L0,\ rp+L0]}(n)=1$ for a p such that $1 \le p \le P\}$

Since it is known that $\Omega = \{n_1<\ldots<n_{Lc}\}$, a grand vector **b** is obtained from the vectors $\hat{\mathbf{h}}_z(n)$, such that:

$$\mathbf{w} = \begin{bmatrix} \hat{\mathbf{h}}_z(n_1) \\ \hat{\mathbf{h}}_z(n_2) \\ \vdots \\ \hat{\mathbf{h}}_z(n_K) \end{bmatrix} \qquad (32)$$

The search for f_0 consists in maximizing the following criterion:

$$\mathrm{Carrier}(f_0)=|\mathbf{w}^H\,\mathbf{c}(\exp(j2\pi f_0 T_e))|^2 \qquad (33)$$

$$\text{where } \mathbf{c}(z_0)=\begin{bmatrix} \mathbf{c}(n_1,z_0) \\ \mathbf{c}(n_2,z_0) \\ \vdots \\ \mathbf{c}(n_K,z_0) \end{bmatrix} \text{ and where } \quad \mathbf{c}(n,z_0)=\begin{bmatrix} z_0^{nI} \\ z_0^{nI+1} \\ \vdots \\ z_0^{nI+(I-1)} \end{bmatrix}$$

The steps of the process suitable for the case of a transmitter with a non-zero frequency are the following:

Step III.a.1: Step I.1 to step I.4 described above in order to obtain the symbol trains $\{\hat{b}_{m-k_j}\}$ associated with the channel vectors $\hat{h}_z(k_j)$.

Step III.a.2: Construction of the vector **w** of equation (32) from the $\hat{h}_z(k_j)$.

Step III.a.3: Maximization of the carrier (f_0) criterion of equation (33) in order to obtain f_0.

Step III.a.4: Application of equation (30) in order to deduce the symbols $\{a_m\}$ from the symbols $\{b_m\}$.

Step III.a.5: Step I.5 to step I.7 described above.

In the case of a transmitter with non-zero frequency and for decorrelated multipaths, the steps are the following:

Step III.b.1: Step II.a.1 to Step II.a.4 described above in order to obtain the vector z(t) of equation (22).

Step III.b.2: Application of the ICA methods [4], [5], [15] and [19] in order to estimate L_c symbol trains $\{\hat{b}_{m,j}\}$ associated with the channel vectors $\hat{h}_{z,j}$.

Step III.b.3: Reordering of the symbol trains $\{\hat{b}_{m,j}\}$ and of the channel vectors $\hat{h}_{z,j}$ applying steps A.1, A.2 and A.3 in order to obtain the symbol trains $\{\hat{b}_{m-k_j}\}$ associated with the channel vectors $\hat{h}_{z,j} = \hat{h}_z(k_j)$.

Step III.b.4: Construction of the vector **w** of equation (32) from the $\hat{h}_z(k_j)$.

Step III.b.5: Maximisation of the carrier (f_0) criterion of equation (33) in order to obtain f_0.

Step III.b.6: Application of equation (30) in order to deduce the symbols $\{a_m\}$ from the symbols $\{b_m\}$.

Step III.b.7: Choice of the symbol train associated with the higher-modulus vector $h_z(i)$, namely, $\{\hat{a}_{m-i}\}$.

Step III.b.8: Step II.a.7 to step II.a.9 described above.

In the case of a transmitter with non-zero frequency and for correlated multipaths, the steps are for example the following:

Step III.c.1: Step II.b.1 to step II.b.6 No. 2.2 in order to obtain the vector z(t) of equation (29).

Step III.c.2: Application of ICA methods [4], [5], [15] and [19] in order to estimate the L_c symbol trains $\{\hat{b}_{m,j}\}$ associated with the channel vectors $\hat{h}_{z,j}$.

Step III.c.3: Reordering of the symbol trains $\{\hat{b}_{m,j}\}$ and of the channel vectors $\hat{\mathbf{h}}_{z,j}$ applying steps A.1, A.2 and A.3 so as to obtain the symbol trains $\{\hat{b}_{m-k_j}\}$ associated with the channel vectors $\hat{\mathbf{h}}_{z,j} = \hat{\mathbf{h}}_z(k_j)$.

Step III.c.4: Construction of the vector **w** of equation (32) from the $\hat{\mathbf{h}}_z(k_j)$.

Step III.c.5: Maximization of the carrier criterion (f_0) of equation (33) in order to obtain f_0.

Step III.c.6: Application of equation (30) for deducing the symbols $\{a_m\}$ from the symbols $\{b_m\}$.

Step III.c.7: Choice among the symbol trains of that one which is associated with the higher-modulus vector $h_z(i)$, namely $\{\hat{a}_{m-i}\}$.

Step III.c.8 : Step II.b.8 to step II.b.10 described above.

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